MILLIMETER-WAVE MICROSTRIP AMPLIFIER USING INDIUM PHOSPHIDE GUNN DIODES

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ABSTRACT

Using small signal data obtained from coaxially mounted low noise Indium Phosphide amplifier diodes, computer optimization has led to a measured reflection stage gain of 4 dB over 28.5 to 38.5 GHz. Higher gain, more narrow band circuits were also derived analytically, and some preliminary results have been obtained with hybrid coupled microstrip amplifiers.

Introduction

For frequencies above 30 GHz, Indium Phosphide Gunn diodes currently provide the best means for low noise wide-bandwidth amplification. Since 1976 the Navy has been supporting Varian Associates' efforts at fabricating both the diodes and circuits towards this end. All of Varian's InP amplifiers utilize the diodes in either coaxial[1] or waveguide[2] cavities which are circulator coupled to separate input from output. In the 26–40 GHz band circulator deficiencies previously have limited the useful bandwidth to about 10 GHz[3]. At frequencies above 40 GHz circulators become even more limiting.

A possibility, which is being explored, is the use of 3 dB quadrature couplers terminated with two identical diode amplifier stages to separate input from output. Since waveguide couplers are fractions of a foot long and microstrip couplers are fractions of an inch long (at these frequencies), the latter are to be utilized.

Ideal 3 dB quadrature couplers have the property that when the coupled ports are terminated in the same impedances, all power not absorbed in these impedances are reflected back to the isolated port. Microstrip 3 dB quadrature couplers have been used at NOSC in the 26.5–40 and 40–60 GHz bands and the present program plans to utilize them together with microstrip reflection amplifier circuits to form wide-band and compact two-port amplifier stages which do not use circulators.

Diode Parameters

In order to computer design a wide-bandwidth resonating structure for any device, it is necessary to know that device's impedance vs. frequency. The impedances used for the designs to follow were obtained from small signal diode characterization in coaxial cavities (by Varian). It would have been preferable to have these measurements made on microstrip and for several months this was attempted (NOSC). The only method the author is aware of for determining such impedances has recently been described by daSilva and McPhun[4]. For each frequency, slotted line data is taken for five structures, all identical except for additional incremental line lengths with one having a short circuit termination instead of an open circuit. A sixth measurement substitutes the device to be measured instead of the of the short.

Results using this method have not been repeatable. This could be attributed to the small physical tolerances allowed for the lines and the position of the microstrip transition on these lines.

Instead, an in-line configuration was chosen for the diode mount. This should approximate coaxial mounting more than if the diode was placed perpendicular to the microstrip line (Figure 1). All circuits to be described in this paper use .010" Duroid (teflon-fiberglass), copper pre-etched to about .004"

Computer Design of Microstrip Reflection Amplifiers

An optimization routine was first written for use with any number of distributed line equalizers (transformers) terminated by the diode impedance, the latter resonated by a single in-line stub (see Figure 1). Perpendicular stubs were avoided since their widths would be fractional wavelengths, making their modeling difficult.

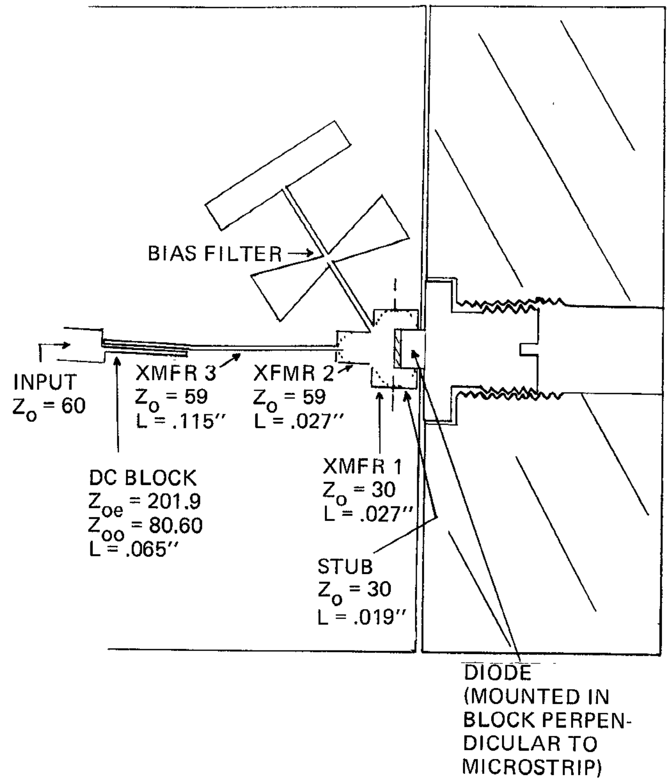

Figure 1. Microstrip circuit optimized for 3 transformers, 1 stub 28, 33, 38 GHz. Dotted outline on XFMRs and Stub show corrections for step discontinuities.

It was hoped that reasonable initial values for all of the variables (transformer and stub lengths and impedances) could be determined by a method based on Getsinger.[5]

In the Getsinger method an equivalent single L, C, -R network is found which behaves like the actual packaged diode. The slope parameter of the tuned circuit is then related to an equivalent low pass prototype. Within bounds set by the first slope parameter, a given number of series and parallel resonators having specific slope parameters (also determined by the low pass prototype) are used to determine the desired gain and bandwidth.

In the Getsinger model, all resonators are inductors and capacitors. Following Cohn,[6] the series LC resonators can be replaced by half-wave distributed lines, and the parallel resonators replaced by other half-wave distributed lines preceded and followed by impedance inverters (quarter-wave distributed lines). Hence, with the exception of the one stub used for resonating the diode, all broadbanding elements consist of transmission lines differing in width and length. Furthermore, a recently disclosed method for compensating step discontinuities[7] can also be used. Instead of steps, each wider line is reduced until it joins the narrower line, the characteristic impedance staying the same because of increased differential edge capacitance. A circuit generated by the method described above can be accurately modeled, even at millimeter wave frequencies. Since the distributed line resonators and impedance inverters are only accurate near the center frequency, a good optimization routine should be able to convert the derived initial values into useful wideband circuits.

To test the optimization routine some arbitrary initial values were first used. An error function, the difference between the average gain with frequency and the desired gain, decreased rapidly. The process was so interesting to watch that a few amplifier models were developed without analytically determining the initial values. Figure 2 shows the result of optimizing three transformers and a stub for 5 dB gain at 28, 33, and 38 GHz. Also shown is the measured result. A quarter wave coupled line matched to the characteristic impedance (60 ohms) of the input line is used as a DC block. Its parameters are used in the network analysis but, in practice, it behaves as an ideal element. Radiation losses and transition losses were not considered in the calculation. The diode used was from the same batch as that for which the impedance data was derived and the same DC bias was used. The measured result was taken with the waveguide to microstrip transition close to the DC block. Transition VSWR can be shown to create relatively large amplitude fluctuations with frequency, and this was found to be true experimentally. Step discontinuity compensation had little effect for this (wide band) case. Considering the high frequencies used, and the chance for error, the above example indicates that measured coaxial diode data can certainly be used for microstrip modeling.

Figure 3a shows the Getsinger model for a two resonator (N=2) amplifier with 8 dB maximum gain and .5 dB ripple. In Figure 3b the LC series circuit is replaced by a half wave resonator, the real diode model is resonated by a stub, and a quarter wave coupled line is used both as a DC block and an impedance transformer. Figure 3c shows the difference between the ideal broadband model and the distributed line equivalent. Figure 4 compares the measured result after the optimization routine was used to slightly increase the stub impedance, greatly decreasing it's width. Step discontinuity compensation was used. When the steps were used without compensation somewhat higher gain resulted. The later amplifier stage was used in the hybrid circuit described below.

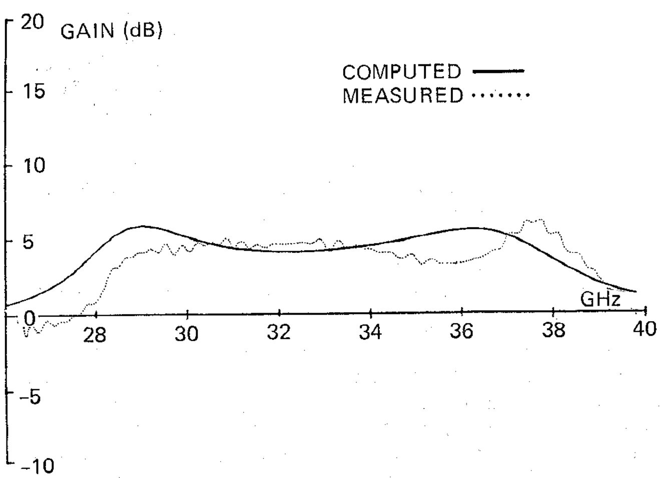

Figure 2. Computed vs. measured gain for 3 XFMR, 1 stub (optimized for 5 dB, 38, 33, 38 GHz).

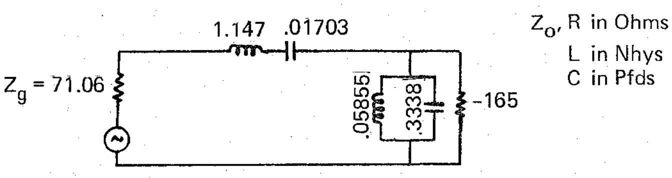

Figure 3a. Getsinger model N=2, 8 dB gain, .5 dB ripple.

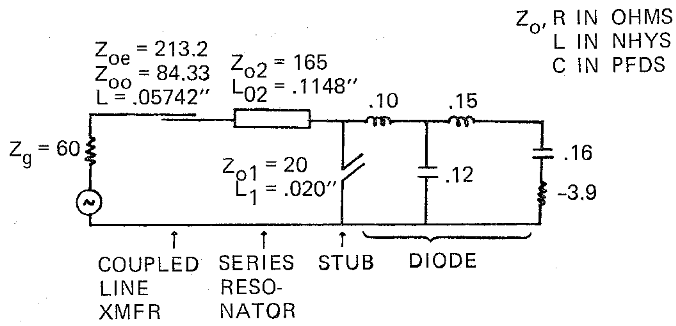

Figure 3b. Practical realization of 3a. Note: Coupled line acts as DC block as well as impedance transformer.

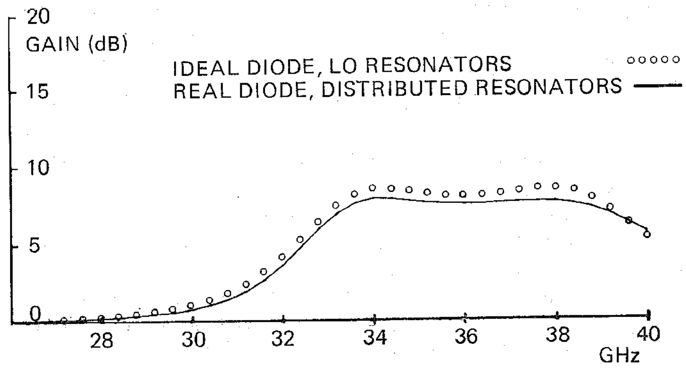

Figure 3c. Comparison of ideal and real circuits.

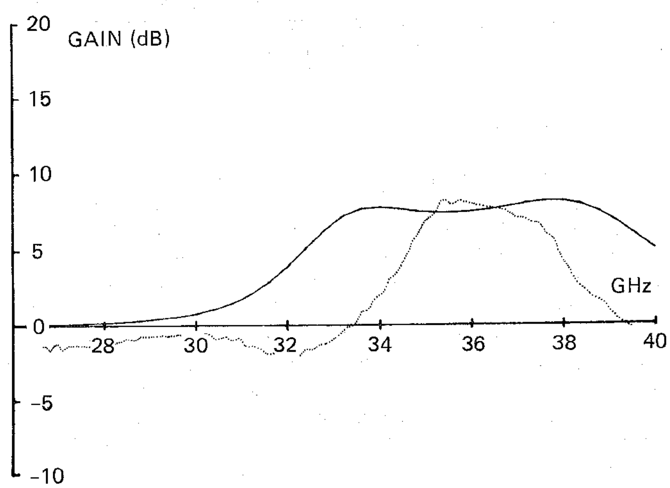

Figure 4. Calculated vs. measured result of circuit optimization based on Figure 3b.
Resonator Z_{02} = 173 ohms, L_2 = .116"
Stub Z_{01} = 22.4 ohms, L_1 = .0223"

Hybrid Coupled Amplifier

All computer data and measurements given so far in this paper are for reflection gain and bandwidth of single diode amplifier stages. To be used as practical amplifiers these single stage devices must either have circulators (as has been done with waveguide), or be combined with other identical states and used as terminations for 3 dB quadrature couplers. Figure 5 shows the result of combining two N=2 non-compensated amplifiers described above. Although clearly short of spectacular, i.e., the bandwidth is limited, a reasonable amount of

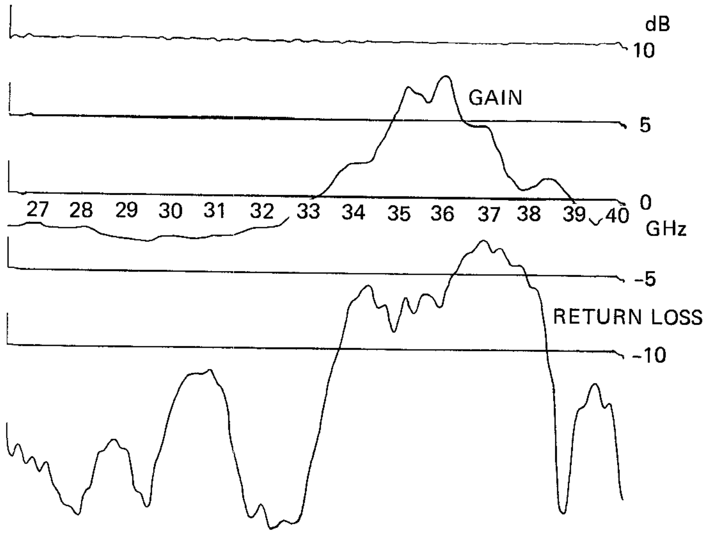

Figure 5. Measured gain and return loss of hybrid coupled amplifier.

isolation existed between input and output. The coupler was a four branch synchronous type with VSWR and isolation measured 20 dB or greater in the amplifier's frequency range. An attempt was made to construct a hybrid coupled amplifier using the low gain 10 GHz bandwidth amplifier but that was not successful.

Conclusions

It has been shown that very broadband microstrip reflection amplifier matching networks can be computer designed and fabricated using impedance data obtained in coaxial cavities. High gain, small bandwidth networks can also be derived analytically. 3 dB microstrip hybrid couplers have been used with dual diode circuits in lieu of circulators, although performance has not been as good as expected.

References

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